AN APPLICATION OF DIGITAL FILTERS FOR RADAR CLUTTER REDUCTION

David Lee Cummings

NAVAL POSTGRADUATE SCHOOL
Monterey, California

THESIS

AN APPLICATION OF DIGITAL FILTERS
FOR RADAR CLUTTER REDUCTION

by

David Lee Cummings

Thesis Advisor: D.B. Hoisington

March 1972

Approved for public release; distribution unlimited.

An Application of Digital Filters

for Radar Clutter Reduction

by

David Lee Cummings
Lieutenant, United States Navy
B.S., University of Colorado, 1966

Submitted in partial fulfillment of the
requirements for the degree of

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

from the

NAVAL POSTGRADUATE SCHOOL
March 1972

ABSTRACT

This thesis presents the design procedures involved in implementing a digital filter for use in a range-gated moving-target-indicator (MTI) radar. The digital MTI filter synthesized on the Xds-9300 computer was a third-order recursive filter with Chebyshev characteristics.

The digital MTI filter was evaluated by constructing the equivalent of a single range channel for use with the AN/UPS-1 air-search radar and comparing its performance to that of the single delay-line canceler of the AN/UPS-1.

TABLE OF CONTENTS

LIST OF FIGURES

I. INTRODUCTION

In the environment in which a pulse radar operates it is frequently difficult to detect moving targets in the presence of a background of clutter. Clutter may be described as any undesirable echo signal from ground, sea, rain, or other unwanted target. It differs from receiver noise in that the echo has a relatively narrow, low-frequency spectrum vice a continuous spectrum. This implies that clutter echoes are correlated from one echo to the next. In many cases the clutter echoes may be many orders of magnitude greater than the echo of a moving target such as a land vehicle or aircraft it is desired to detect.

The basic criteria used for distinguishing between fixed and moving targets is the doppler shift in frequency of the moving target. The doppler frequency shift is given by the relationship

$$f_d = \frac{2 v_r}{\lambda}$$

where v_r is the radial component of velocity of the target relative to the radar and λ is the wavelength of the transmitted signal.

The presence of various frequencies is the basis for the use of a filter that passes echoes with a doppler shift but rejects clutter which has zero or small doppler shift.

In the time domain the moving target echoes appear in the video amplifier of a coherent radar with a phase detector as pulses continuously varying in amplitude at the doppler frequency as sampled at the pulse repetition rate. Fixed targets, on the other hand, have a constant pulse amplitude which may be either positive or negative.

An analog or digital filter can function to separate moving targets from fixed clutter in an MTI radar.

A. ANALOG IMPLEMENTATION

In the past the most common method of constructing an analog MTI filter was by the use of ultrasonic delay lines. Figure 1 shows a simplified block diagram of a single delay-line canceler incorporated into a basic radar. The output of the phase detector is applied to two channels. One channel contains the delay line where the information is delayed one interpulse period. The output of the delay line is subtracted from the output of an undelayed channel. Thus under ideal conditions successive fixed-target echo pulses will cancel since they are of equal amplitude whereas the differing pulse-to-pulse amplitudes of the moving targets will leave some value of residue upon subtraction.

A portion of the frequency response of the single delay-line canceler is shown in figure 2. This characteristic is periodic in the frequency domain at multiples of the pulse repetition frequency. This response would be adequate if all clutter was stationary with respect to the antenna. In actuality the clutter has a finite frequency spectrum caused by echoes from rain, sea, and vegetation fluctuating both in phase and amplitude with time as detailed in Ref. 1. Other fluctuations are introduced by the rotational scanning of the antenna. A portion of the resultant clutter spectrum in a typical case and the desired clutter rejection filter spectrum are shown in figure 3. This spectrum repeats at multiples of the pulse repetition frequency and extends to approximately the reciprocal of the transmitted pulse width.

To improve the frequency response of the MTI canceler, two single delay-line cancelers may be cascaded. The resultant frequency response of a double delay-line canceler is shown in figure 4.

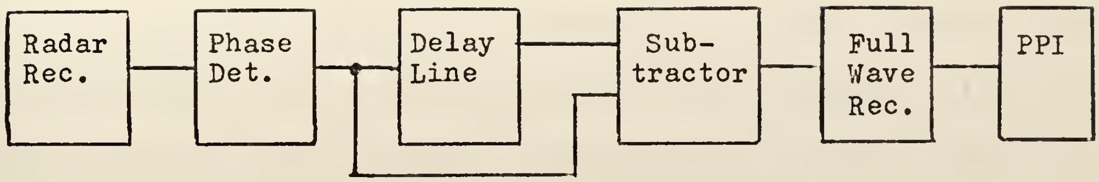


Figure 1 Single delay-line block diagram

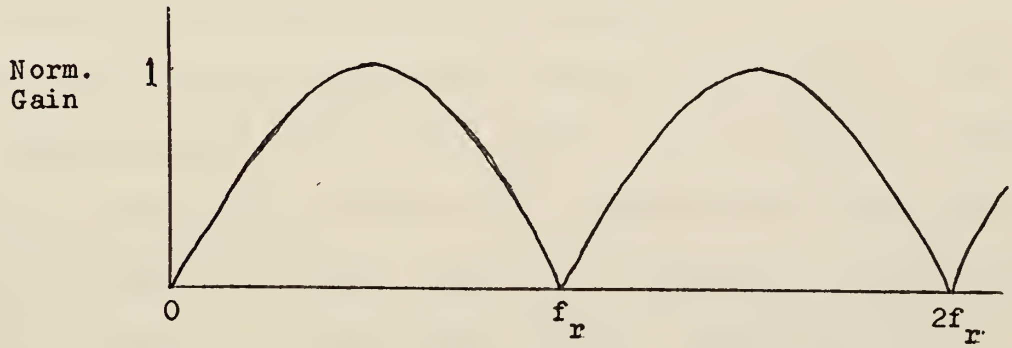


Figure 2 Frequency response of single delay line

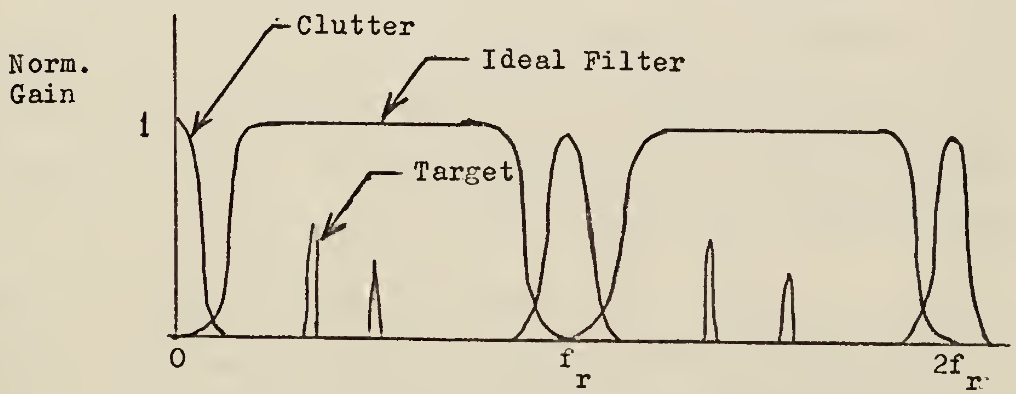


Figure 3 Frequency spectrum of clutter and ideal filter

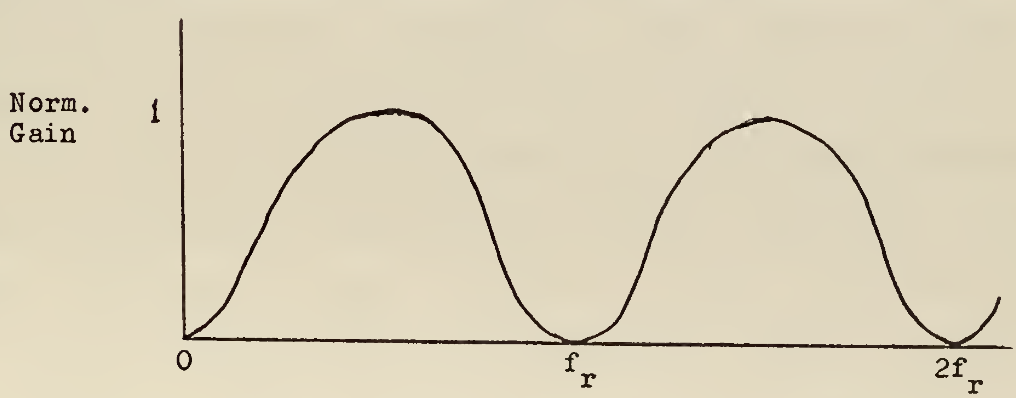


Figure 4 Frequency response of double delay line

The disadvantages of acoustical delay lines arise primarily from their physical characteristics. Individual lines have insertion losses as high as 55 dB and the resultant amplification required is hard to balance between the two channels. Other problems include changes in delay due to temperature variations and introduction of unwanted secondary signals through multiple reflections.

An analog MTI system may also be implemented by the use of active elements. This process requires that targets be separated on a time basis. Bipolar video is applied to a number of range channels arranged in parallel, each of which transmits only a range interval comparable to the pulse width of the radar. Successive return echoes from the same target are processed by the same range channel. Sampling the video signal at the pulse repetition frequency limits the frequency response of the resulting signal to half the pusle repetition frequency. Within each range channel the moving targets can be distinguished from the fixed targets by means of a clutter-rejection filter which eliminates components near zero frequency (clutter), but transmits higher frequency components associated with moving targets.

Building a MTI employing range gates and filters requires the construction of many similar channels out to the maximum range where clutter may be encountered. For example if an MTI capability out to 70 nautical miles was desired for the AN/UPS-1 with its 1.4 micro-second pulse width, the range-gate method would require about 620 channels.

The range-gate-and-filter method provides many improvements over the previous delay-line MTI systems. The first of these is that more nearly ideal clutter rejection filters can be implemented, namely two, three, or four-pole filters vice the one pole equivalent of the single delay-line.

The failure of one range gate does not render the system totally inoperative as would a single failure in a delay-line canceler. In addition the passband of the filter can be altered by changing certain components within the filter to match the filter to changing clutter spectra.

Previously the major reason for not using the range-gate-and-filter technique was the large number of channels required and hence the extensive number of electronic elements necessary. With advances in the field of integrated circuits, the construction of a system in a reasonably sized unit is feasible as shown in Ref. 2.

B. DIGITAL IMPLEMENTATION

The implementation of a digital filter for use in a range-gated MTI system has been made possible by the rapid advances of recent years in the field of digital components. The flexibility, speed of operation, and data-handling capability of the digital hardware makes it invaluable in the processing of radar video signals. Concurrent with the improvements in hardware there has been an equally impressive development of the theories and methods of digital signal processing.

The basic scheme is the same as that of a range-gated analog MTI system, but with the A/D converter and digital storage replacing the analog cancellation filter and associated circuitry.

The digital MTI system possesses many advantages over any delay-line system in that it is inherently more flexible, stable, and reliable. From an economic point of view it compares quite favorably to a range-gated analog MTI.

The remainder of this paper develops the implementation of a digital MTI and the construction and evaluation of a single range channel for use with the AN/UPS-1 air search radar.

II. DIGITAL MOVING TARGET INDICATOR

A digital moving target indicator for a air-search radar may be realized by the correct arrangement and operation of digital hardware. A special processor rather than a general-purpose digital computer is required due to the necessity for a rapid clock rate of from one to several megahertz. This requirement is brought about by the narrow pulse width of the radar.

Implicit in the idea of digital signal processing is the conversion of a analog video signal to a series of digital values. These digital values may then be stored, retrieved, weighted, and combined in a manner prescribed by a transfer function to accomplish a desired result, in this case a moving target indicator (MTI) filter.

The basic principles of applying digital filters for moving target indicators is given in Refs. 3 - 5. A block diagram of a basic first-order digital MTI filter is shown in figure 5.

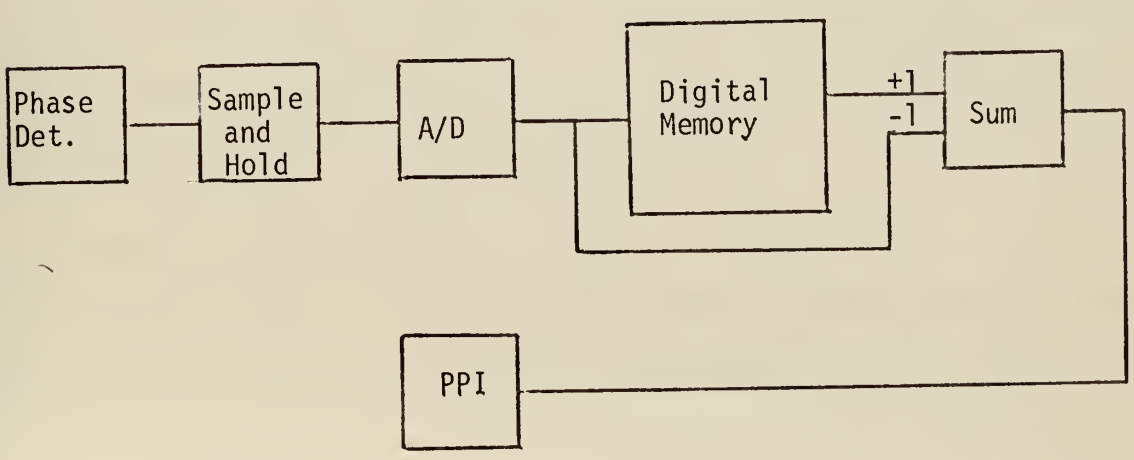


Figure 5 - Basic digital MTI System

By implementing the MTI filter with high-speed digital components, only one sample-and-hold circuit and one A/D converter are required along with a memory of sufficient size to store the digital values of the echoes arriving from the successive range channels and the necessary arithmetic circuits (multiplier and summer).

A number of identical parallel channels will be necessary if digital components with clock speeds less than those used above are employed. The increase in the number of channels being directly related to the decrease in clock rate.

A. SAMPLE AND HOLD

The sample-and-hold circuit performs the operation of sampling the bipolar video output of the radar receiver's phase detector. To prevent any loss of information the sampling is done at discrete range intervals corresponding to the radar's pulse width. A typical sample width is usually in the range of 10 nanoseconds to the sampling period and for the AN/UPS-1 radar the sampling would take place with a period of 1.4 microseconds.

The sampling operation requires a stable and accurate clock to ensure that the successive sampled values are selected at the same time after the transmitter pulse as were the previous values.

B. ANALOG-TO-DIGITAL CONVERTER

The analog-to-digital (A-D) converter forms a digital word by encoding a sequence of binary digits equivalent to the sampled value. The accuracy to which the digital word correctly represents the analog value is limited by the length of the binary sequence. The effect of the finite word length is known as the quantization error of the input. This effect is the subject of many papers such as Ref. 6.

The number of bits that make up a word is one factor that will determine the attainable clutter suppression of the digital filter. Too long a word length tends to have the last bit reflect internal noise rather than the input signal. The use of 10 bits results in good conversion for radar video as it provides 512 levels in a positive or negative sense, a total of 1024 different quantization levels.

C. MULTIPLIERS

The multipliers that appear in second and higher order forms of the digital filter must also be represented by a discrete word length. It is advantageous to use short word lengths (3 or 4 bits) so that the time necessary for the operation is not excessive. This results in some restriction on the value of the filter coefficients and leads to slight deviations from the desired filter characteristics as noted in Ref. 3.

D. DIGITAL MEMORY

For the MTI filter to have a capability out to the maximum range at which clutter may be encountered requires that the output of the A-D converter be processed in the digital filter in a time-multiplexed manner.

The digital value of each successive range channel is sent to the digital memory for the first-order system shown in figure 5. After one interpulse period the value is retrieved in time sequence and digitally subtracted from the current digital value of the A-D converter for that range channel. The storage of the digital value for the interpulse period is equivalent to the delay line of the delay-line canceler.

The implementation of the digital memory could be accomplished by a magnetic core, integrated shift register, or a integrated scratch-pad

memory to be determined by a study of hardware characteristics at the time a particular system is implemented.

The number of bits of memory required is determined in the following manner as outlined in Ref. 4. The maximum range (R) in nautical miles to which clutter may be encountered and the sample rate (T) in microseconds determine the number of range intervals or digital words (W).

$$W = \frac{12.3(R)}{T}$$

where the constant 12.3 is the number of microseconds per nautical mile. The total number of digital bits (TB) of storage required depends on the number of bits per word (B) in the A-D converter and the order of the filter (N).

$$TB = (B)(W)(N)$$

For example, a digital MTI system for the AN/UPS-1 radar out to a range of 70 nautical miles employing a ten bit per word conversion rate and a third order digital filter would require

$$W = 615 \text{ words}$$

$$\text{and} \quad TB = 18{,}450 \text{ bits of memory}$$

After the subtraction process the successive digital values are converted to an analog voltage in the D-A converter. By using a D-A converter of one less bit than the A-D converter the output analog voltage is unipolar for use in a plan-position indicator.

E. DIGITAL FILTER

The type of filter shown in figure 5 is a first-order, non-recursive filter. Its frequency response is equivalent to that of the single

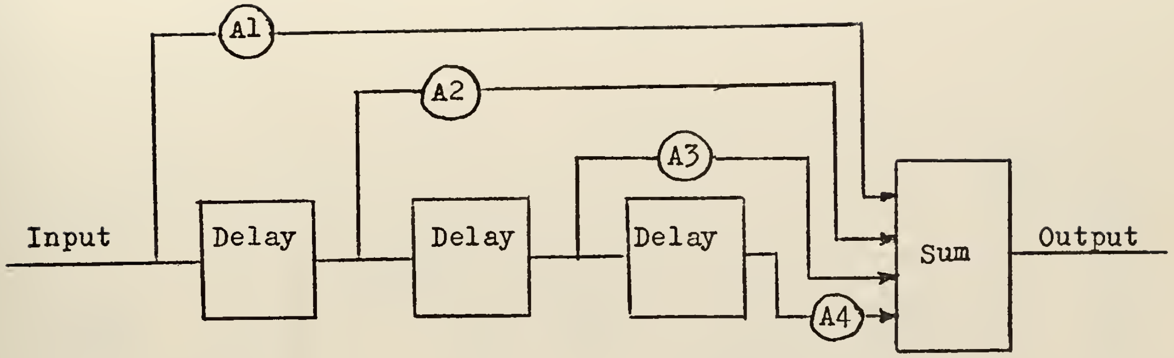
delay-line canceler. An improvement in clutter-rejection capability may be gained by using a second or third-order digital filter.

Digital filters may be broadly classed as either recursive or non-recursive or may be a combination of both types. The basic form of a third-order non-recursive digital filter and its transfer function is shown in figure 6. Its impulse response is of finite duration as it can be noted that the present value of output depends only on the present and two past values of the input. The zeros of the z plane transfer function are real and/or complex conjugate pairs while the three poles are all at the origin.

The form of the third-order recursive digital filter and its transfer function is shown in figure 7. Its impulse response is of infinite duration as here the output is a linear combination of the input and past outputs. The zeros are at the origin and the three poles are real and/or complex conjugate pairs.

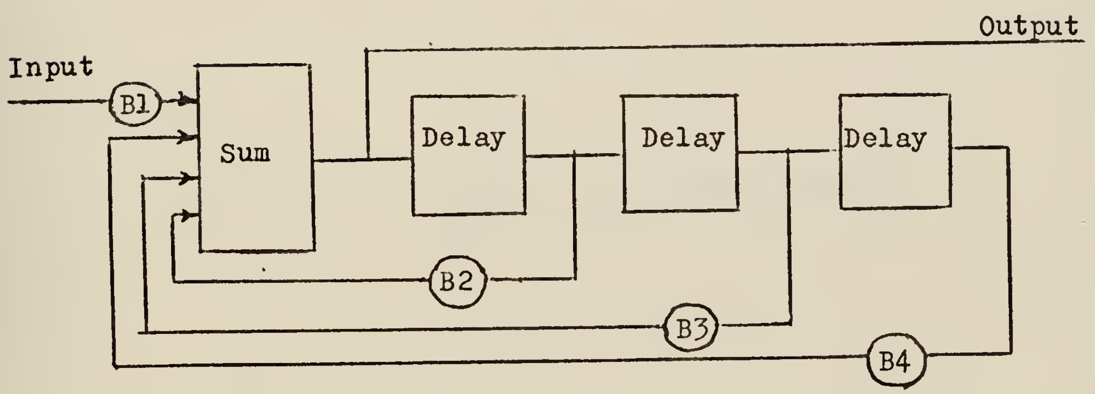
In order to implement a transfer function that is the ratio of two equal-order polynomials a combination of recursive and non-recursive filters may be used in the canonical configuration. The form of a third-order canonical filter and its transfer function is shown in figure 8. There are several other basic digital network configurations which may be used to yield the same z-domain transfer function; the direct form, the parallel form, the cascade form, and the coupled form.

These various forms are shown and explained in Ref. 7 along with the suggestion that it is never advisable to implement third or greater order filters in the direct or canonic form due to the possibility that the filter will be unstable resulting from quantization error in the pole and zero placement. The suggested remedy is to synthesize all high-order



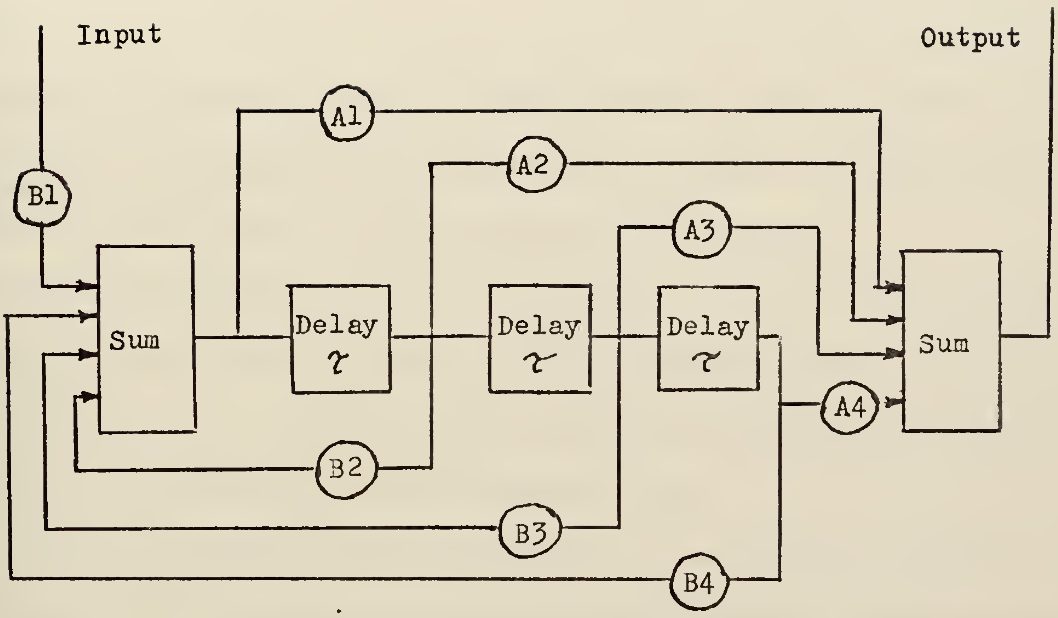
$$H(Z) = \frac{A1\ Z^3 + A2\ Z^2 + A3\ Z + A4}{Z^3}$$

Figure 6 Third-order non-recursive filter



$$H(Z) = \frac{Z^3}{B1\ Z^3 - B2\ Z^2 - B3\ Z - B4}$$

Figure 7 Third-order recursive filter



$$H(Z) = \frac{A1\ Z^3 + A2\ Z^2 + A3\ Z + A4}{B1\ Z^3 - B2\ Z^2 - B3\ Z - B4}$$

Figure 8 Third-order canconical configuration

filters by parallel and cascade connections of first or second order filters, or to be extremely careful in the analysis and design of higher-order filters.

The non-recursive digital filter must be of a greater order than the recursive filter to obtain the same sharp transitions in the frequency response. Since the recursive filter requires fewer terms than the non-recursive to accomplish the same frequency response characteristic it is more practical from a real-time implementation view.

The goal of this digital filter design is to derive the transfer function with optimum clutter rejection characteristics. The frequency response of the filter of a given order is determined by the coefficients of the feed-forward and feed-back loops.

These coefficients may be determined in the following manner. First a analog filter having the required frequency-response characteristic is chosen. The x-plane transfer function of this filter is mapped into the z plane by the use of a certain transformation. The resulting transfer function, in the form of a ratio of polynomials in z, is implemented as a digital filter in one of the configurations previously discussed.

This procedure lends itself to making use of the numerous design techniques for analog filters which have been developed over the past years and discussed in Refs. 8 and 9. These designs employ the use of the well-tabulated standard filter forms referred to as Butterworth, Chebyshev, Bessel, and others. Each form has certain characteristics which determine the most appropriate use for a given filter.

Once the desired frequency-domain transfer function, H(S), has been specified, the bilinear z transform is used to determine the corresponding digital transfer function, H(Z).

The bilinear z transform is stated by the following equation

$$s = \Omega \frac{z+1}{z-1} \qquad \text{where } \Omega = \tan\frac{\phi}{2} = \tan\frac{\pi f_c}{PRF}$$

and is a one-to-one mapping of the s plane into the z plane. If H(S) specifies a low-pass filter function then the bilinear z transform ensures that H(Z) will be a low-pass transfer function. The filter response in any digital system is periodic in the frequency domain at multiples of the sampling frequency. In the digital MTI filter the sampling occurs at the pulse repetition frequency which results in the frequency response being symmetrical about the midpoint between adjacent PRF lines. The effect of sampling at the PRF is the synthesis of a comb filter which can be used for a moving target indicator.

The scale adjusting parameter Ω allows for selection of the location of the filter corner frequency (f_c) at a point $\phi/2\pi$ times the pulse repetition frequency (PRF). In essence Ω determines the width of the filter passband.

It is also noted that this type of digital filter cannot distinguish between a positive or negative doppler shift.

III. A DIGITAL MTI FOR THE AN/UPS-1 RADAR

The previous section described the general characteristics and components of a digital MTI system. These ideas are applied to developing a digital MTI for the AN/UPS-1 radar by using an available general-purpose computer. This air-search radar has a operating frequency of about 1300 megahertz, a pulse repetition frequency (PRF) of 800 hertz, and a transmitted pulse width of 1.4 microseconds. Its internal single delay-line canceler was used for comparison of MTI performance.

Two limitations were imposed on the system design due to the characteristics of the equipments involved. First the equivalent of only one of the many range channels found in the analog range-gates and filters method was implemented because the cycle time of the computer available for use was greater than the AN/UPS-1 radar's pulse width. This is not a major limitation for this demonstration because in an actual system the successive digital values representing the echoes existing in the parallel range channels of the analog range-gates and filters method are processed sequentially by the same arithmetic circuits (multipliers and summers). The evaluation of the equivalent of one range channel is considered valid.

The other limitation resulted in not being able to perform the total operation in real time since the radar system is not located adjacent to the computer facility. The data developed from the radar was recorded on a magnetic tape recorder and then moved to the computer facility for the signal processing.

A block diagram of the system is shown in figure 9.

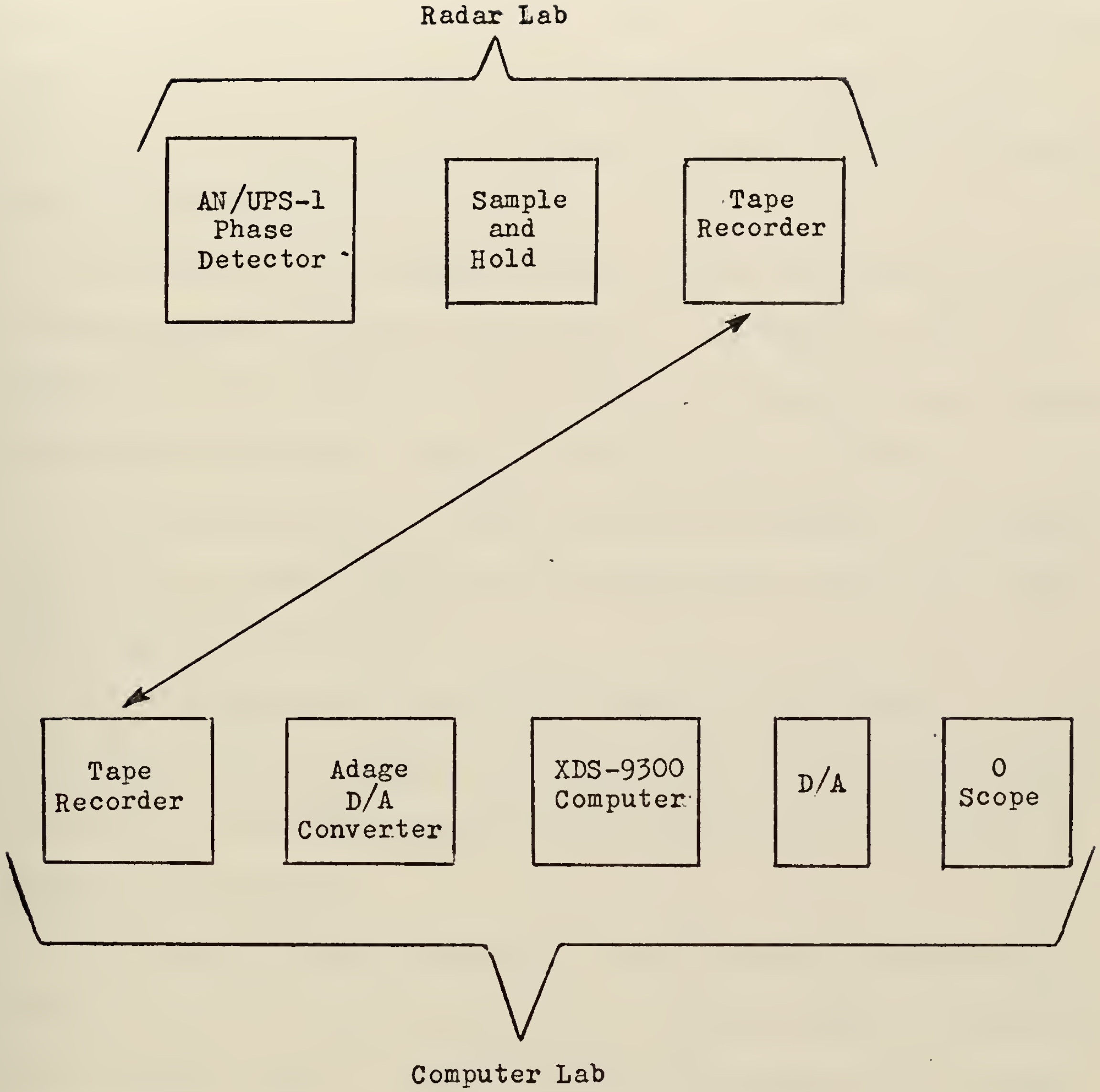


Figure 9 Block diagram of digital MTI system

A. SAMPLE-AND-HOLD CIRCUIT

The input to the sample-and-hold circuit is the bipolar video output of the phase detector of the AN/UPS-1 radar. The circuit is required to sample the same 1.4 microsecond range interval at successive pulse repetition periods and to hold the sampled value for 1.25 milliseconds, the pulse repetition period.

The inclusion of the sample-and-hold circuit also enables one to use a narrow-bandwidth magnetic tape recorder. A bandwidth of about one megahertz would be required if the video were recorded without sampling. This would necessitate a wide-band state-of-the-art recorder.

The circuit as shown in figure 10 was used previously in an analog MTI system that used range-gates and active analog filters for clutter rejection as noted in Ref. 10.

The Darlington-pair input to the range gate was required to provide the current drive to charge the holding capacitor to the maximum value of the input video during the time the gating pulse from the pulse generator was present.

The holding circuit is basically a boxcar generator implemented by the use of FET's. The circuit provides a constant output voltage equal to the sampled input for a period of 1.25 milliseconds along with the necessary sharp rise and fall transitions that occur at each new sample time.

The range interval covered by the channel is controlled by a variable-delay pulse generator triggered by the AN/UPS-1 system trigger. The channel then processes the target signals, clutter, and any noise that is present during the 1.4 microsecond sampling interval. The bandwidth of the sampling circuit is adequate to prevent collapsing loss. Another

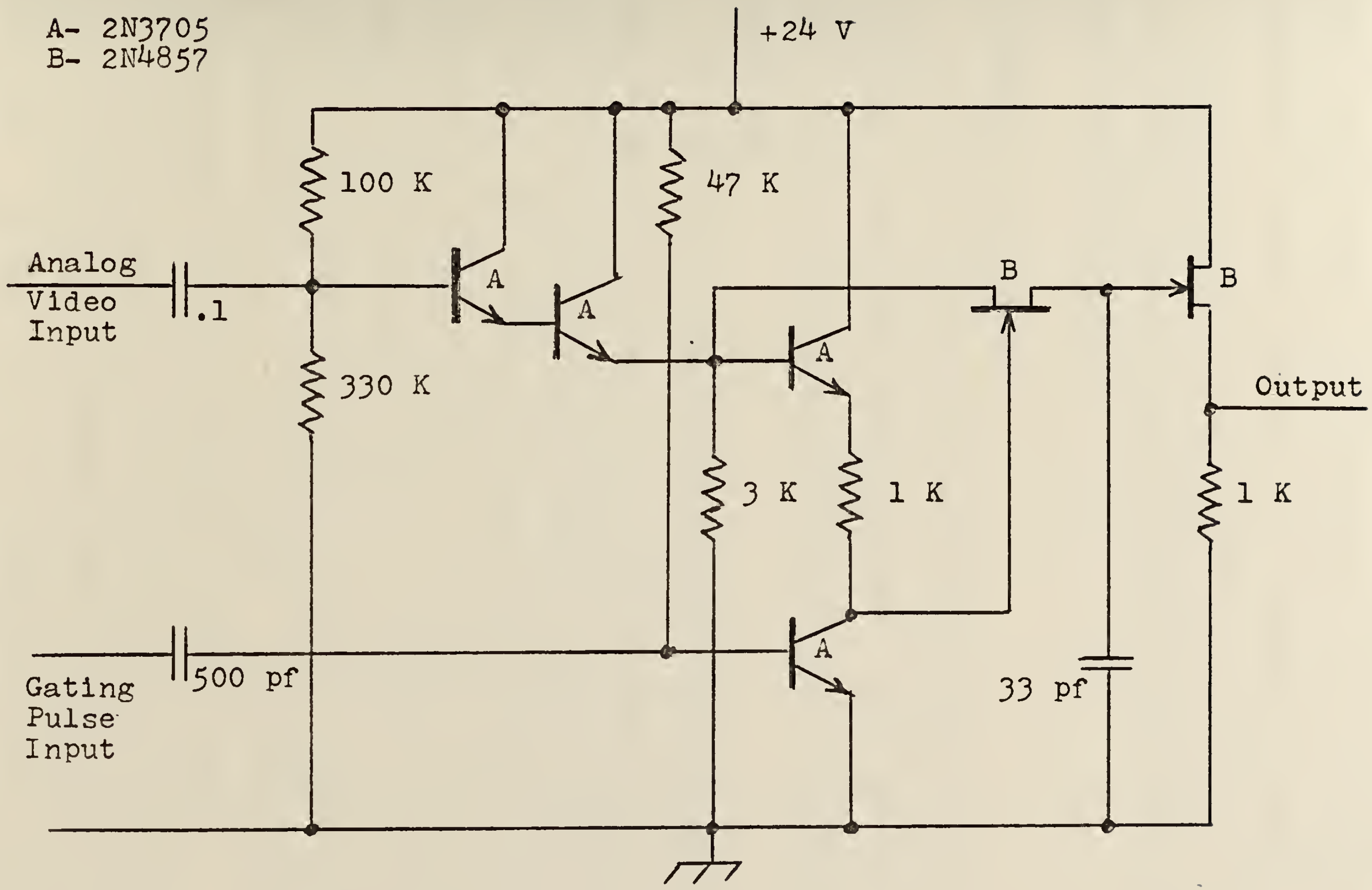


Figure 10 Sample-and-hold circuit

characteristic of the sampling process is that signals associated with all doppler frequencies except multiples of the pulse repetition frequency (blind speeds) are in the output, but are converted to a fundamental frequency of not more than one half the sampling frequency.

Following the sample-and-hold operation the resultant audio waveform was recorded on magnetic tape for transfer to the computer facility for digital signal processing.

B. MAGNETIC TAPE RECORDER

A portable seven-track instrumentation recorder was used to record the output of the sample-and-hold circuit. It was operated in the FM recording mode so that frequencies from dc up to 2.5 kilohertz could be retained.

FM recording is accomplished by deviating the carrier frequency in response to the amplitude of a data signal. A dc signal of one polarity increases the carrier frequency while the opposite polarity decreases it. An ac signal alternately increases and decreases the carrier above and below its center frequency at a rate equal to the data signal frequency. In the reproduction process any amplitude instability in the carrier is eliminated by limiting, and the data signal is reconstructed by detecting the rate of zero crossings. The residual carrier signal and out-of-band noise are removed by a low-pass filter.

A problem associated with FM recording is its sensitivity to tape speed fluctuations or flutter since these produce unwanted frequency modulation. The advanatages of FM recording are that it gives reasonably good amplitude stability, response down to zero frequency, and good linearity for both dc and ac signals. To gain these advantages over the

direct recording method the frequency response for a given tape speed is greatly reduced, the recorder is more complex and expensive, and there is a need for a more constant tape speed. More information on magnetic tape recording is available in Ref. 11.

C. DIGITAL EQUIPMENT

The clutter rejection filter of the digital MTI system was implemented on an available hybrid-computer system consisting of the following main subsystems:

1. Comcor Ci-5000 analog computer
2. Adage VT-13 14 bit A/D converter interfaced between 1. and 3.
3. Xerox Data Systems XDS-9300 digital computer

Since the above general-purpose computer system cannot be hard wired for a specific user the third-order recursive filter was synthesized by a digital computer program written in the FORTRAN programming language and used in conjunction with a subroutine that provides a more rapid D/A and A/D conversion than is normally available. This subroutine was written in the symbolic language, METASYMBOL, which results in a machine-language program being generated in the XDS-9300 computer. The use of this subroutine allowed the computer to execute the FORTRAN program for the digital filter at a speed fast enough to process in real time the equivalent of the one range channel that was implemented.

The sampled analog signal from the magnetic tape recorder was inputed to the patchboard of the analog computer. It was then fed directly to a trunk line connected to one of the input channels of the A/D converter. The necessity for using the analog computer occurs only because of the particular hard wired hybrid configuration found at the facility used for

implementing the filter where all inputs to the A/D converter must go through the analog patchboard. The operation of the A/D converter is controlled by the computer program. When directed by the program the analog value is converted to a digital value as previously discussed and stored in a specified location in memory.

Once the conversion has taken place the computer then performs the arithmetic operations as specified by the program. Upon the completion of the sequence of multiplications, additions, and shifting of register values the derived output value for that sample period, 1.25 milliseconds, is simultaneously stored for future use and made available to the D/A converter for displaying the filter output. This is done by transferring the output of the D/A converter via a trunkline to the analog patchboard where it may be viewed on a oscilloscope.

The process described above is repeated every 1.25 milliseconds as controlled by a accurate and stable external signal generator in conjunction with test lines referred to in the computer program.

The computer program is based upon the digital transfer function, H(Z), of the required clutter rejection filter. The filter design is described in the following section.

D. CLUTTER REJECTION FILTER DESIGN

It was shown in figure 2 that the ideal clutter rejection filter had characteristics of an ideal bandpass comb filter with the passband excluding the band of frequencies around each PRF line where the maximum clutter is present.

When the ideal filter rejects clutter it also eliminates any moving target whose doppler frequency is within the excluded band of frequencies around each PRF line. These so called blind speeds are given by

$$v_n = \frac{n\lambda f_r}{2} \qquad n = 1,2,3,\ldots$$

where v_n is the n^{th} blind speed, λ is the wavelength of the radar carrier, and f_r is the pulse repetition frequency.

For the AN/UPS-1 radar the first blind speed is 181 knots. Thus if the ideal filter rejected clutter with doppler frequency shifts from zero to eight percent of the PRF frequency the resultant blind speeds, in knots, would be

0 - 15, 166 - 196, 347 - 377, etc.

The width of the clutter rejection bands should be neither too narrow nor too wide. If too narrow much clutter will not be eliminated. If too wide many moving targets will be eliminated.

It is apparent from the above that the selection of the proper clutter rejection band is a compromise, and it would be desirable to be able to alter the rejection width so that the optimum filter could be used for the particular clutter conditions existing at any given time.

A choice must be made of the type of filter that will provide the desired frequency response. Two of the standard types are the Butterworth filter and the Chebyshev filter. The third order frequency responses of these filters are shown in figure 11 as obtained from Ref. 12.

It can be seen that the Chebyshev filter with 0.5 dB ripple in the passband provides a more uniform transmission of frequencies and a sharper rate of cutoff.

The frequency-domain transfer function of a Chebyshev filter may be specified by stating the maximum amount of ripple in the passband

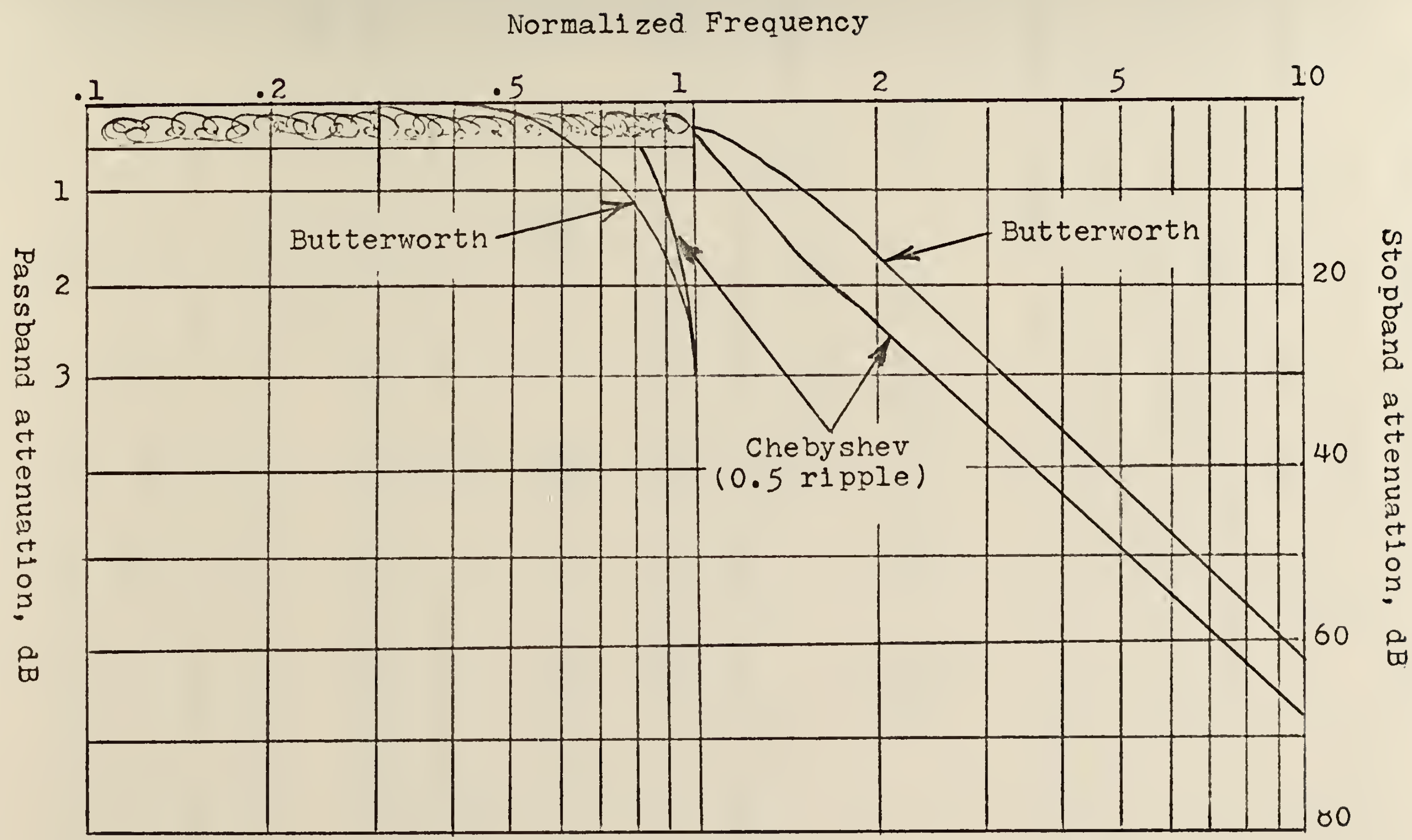


Figure 11 Frequency response of Butterworth and Chebyshev filters

and either the order of the function or the rate of cutoff outside the passband.

The clutter rejection filter for the digital MTI for the AN/UPS-1 radar was chosen to be of the Chebyshev type. A third order filter was synthesized because the time required to handle the various arithmetic operations and register changes could be adequately done by the computer in the 1.25 milliseconds between transmitted pulses.

The frequency-domain transfer function, H(S), of a third order Chebyshev low-pass filter with 0.5 dB ripple in the passband is

$$H(S) = \frac{1}{s^3 + 1.2529s^2 + 1.5349s + 0.7157}$$

as given in the tables of Ref. 13.

The one-to-one mapping of the s-plane transfer function, H(S) into the z plane is done by the bilinear z transform

$$s = \Omega \frac{z+1}{z-1} \qquad \text{where} \quad \Omega = \tan \frac{\phi}{2} = \tan \frac{\pi f_c}{PRF}$$

As noted in the previous section the scale adjusting parameter Ω controls the location of the filter corner frequency (f_c) as it is mapped into the z plane.

It can be noted that if Ω is unity, ϕ is 90 degrees which corresponds to a passband from 200 to 600 hertz for the AN/UPS-1 radar. If ϕ equals 60 degrees then the passband will extend from 133 to 667 hertz. Likewise if ϕ is reduced to 30 degrees the passband will expand to cover 67 to 733 hertz.

The above two values of ϕ result in scale adjusting parameters of

$$\Omega = \tan \frac{60^\circ}{2} = 0.577$$

and $$\Omega = \tan \frac{30^\circ}{2} = .0268$$

Using these values and the bilinear z transform the mapping of H(S) to H(Z) results in the following digital transfer functions

$$H(Z) = \frac{.452\ (z - 1)^3}{z^3 - .923z^2 + .643z - .025} \qquad \phi = 60^\circ$$

$$H(Z) = \frac{.805\ (z - 1)^3}{z^3 - 1.94z^2 + 1.37z - .3} \qquad \phi = 30^\circ$$

These digital transfer functions were programmed on the digital computer. Computer Program 1 and Computer Program 2 were used to reject clutter that was recorded on the magnetic tape when the AN/UPS-1 radar was operated. The frequency responses of these filters is shown in figure 12. Their performance is evaluated in the following section.

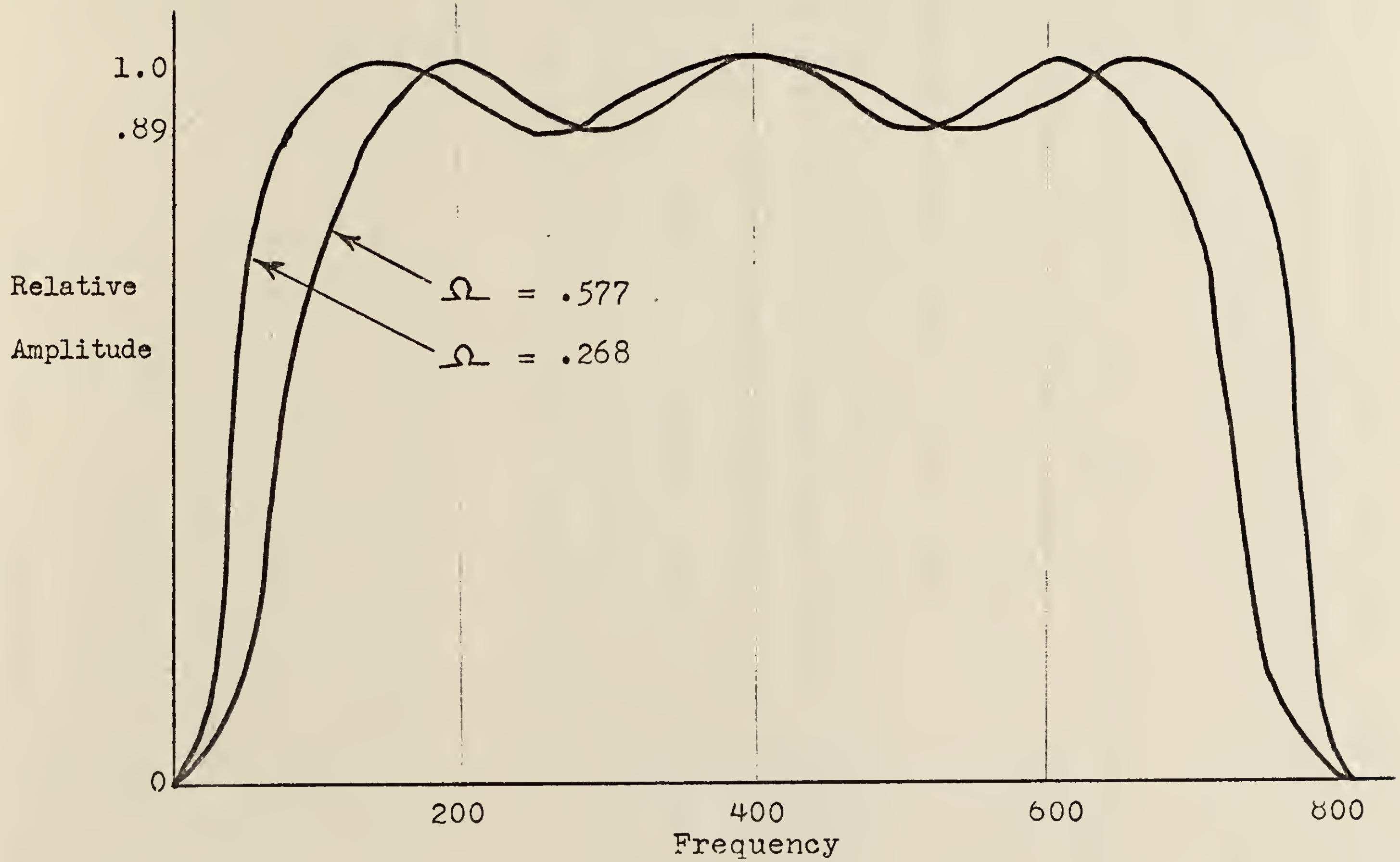


Figure 12 Digital transfer function frequency responses

IV. EVALUATION OF THE DIGITAL MTI USED WITH THE AN/UPS-1

The evaluation of the digital moving-target indicator was done by comparing its performance with that of the single delay-line canceler of the AN/UPS-1 radar. The two parameters which were measured and formed the basis for a comparison were minimum discernible signal and subclutter visibility.

The minimum discernible signal (MDS) is the weakest echo that the radar receiver is able to detect and process. It is limited by both the receiver and atmospheric noise that occurs within the same part of the frequency spectrum as does the echo signal.

The subclutter visibility (SCV) is a measure of the moving-target indicator's ability to detect moving-target signals superimposed on clutter signals. It is defined as the gain in signal-to-clutter power ratio produced by the MTI. For example, a SCV of 20 dB implies that a moving target can be detected in the presence of clutter even though the clutter echo power is 100 times the target echo power.

A. PROCEDURE

The tracking of a real target was tried but could not be accomplished for any length of time because the aircraft moved too rapidly in both range and azimuth to manually locate the range cell at the proper position.

By injecting the equivalent of a moving target from a signal generator into a directional coupler located in the waveguide of the radar accurate measurements could be made of the minimum discernible signal and subclutter visibility. Figure 13 shows a block diagram of the equipment set-up.

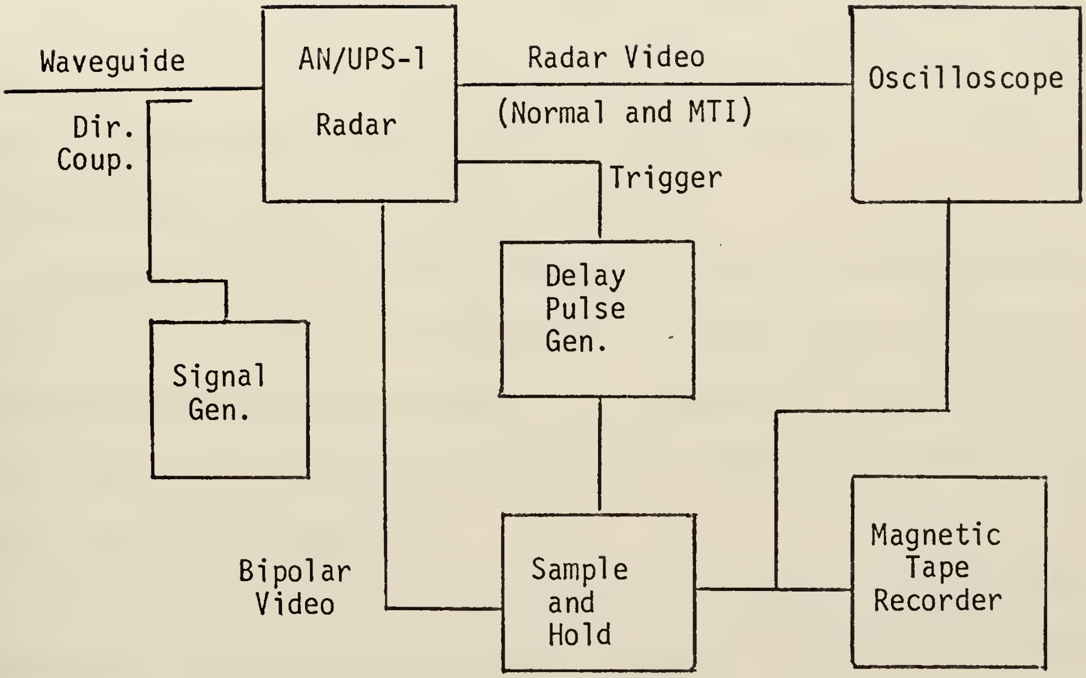


Figure 13 Equipment set-up for measurement of MDS and SCV

For a given input signal, the output of the AN/UPS-1 radar's delay-line canceler could be observed on its own A-scope and plan-position indicator and on an external oscilloscope while the output of the sample-and-hold circuit, the only element of the digital MTI system in the radar lab, was recorded on magnetic tape.

The recorder had seven channels and was used in the evaluation phase in the following manner. A recording of the output of the sample-and-hold circuit for a given input signal power was made on one channel while noting the level of output of the delay-line canceler on the oscilloscope.

This process was then repeated six times on the other channels, each time with a discrete reduction in input signal power, before the recorder was moved to the computer lab for the remaining signal processing.

In the computer laboratory the several signals were processed in turn until no discernible signal was noted on the oscilloscope. This constituted the MDS of the digital MTI for comparison with the MDS obtained with the radar's delay-line canceler.

Although the actual minimum discernible signal for each system was determined, it is their relationship to each other which is important because the absolute value depends on the particular state of tune of the radar at the time of measurement, and to some extent upon the individual making the comparison.

The measurement of subclutter visibility was done with the radar transmitter operating and the antenna stopped in a direction of much ground clutter. A coherent moving target signal was simulated by superimposing the output of an audio signal generator on the clutter at the input to the sample-and-hold circuit. In the delay-line canceler the measurement was made by moving the signal from the external signal generator to a area of high clutter.

B. RESULTS

The minimum discernible signal (MDS) was measured in the absence of clutter. For the digital MTI and MDS was the same as that measured for the normal non MTI video of the radar, -105 dBm. The MDS of the delay-line canceler was -96 dBm. Thus the digital MTI has 9 dB greater sensitivity than the delay-line canceler due primarily to improved frequency response characteristics and lower internal noise.

The subclutter visibility was measured in the manner previously described. The SCV of the delay-line canceler was 21 dB while that of the digital filter was 25 dB, a 4 dB improvement.

V. CONCLUSIONS

This thesis was motivated by an interest in digital signal processing and the resultant advantages these techniques can bring when applied to a specific problem, the distinguishing of a moving radar target in an environment of excessive clutter.

Although the digital equipment that was used in this case was a general-purpose computer rather than a special processor that would be used in an actual application, the techniques and theory are the same.

As the rapid advances in digital components continues it becomes more and more feasible from an economic as well as a performance point of view to consider employing their advantages, namely stability, reliability, and flexibility, to the digital processing radar video.

An example of the flexibility of digital processing is the relative ease with which the coefficients of the realized filter can be changed to reflect variations in clutter conditions in the operating area. Another use to which digital filters are readily adaptable occurs in the more advanced radars which have a variable pulse repetition frequency to eliminate the occurance of blind speeds.

The general conclusion to be made upon completion of the design and testing of the digital filter was that the use of digital signal processing of radar video can provide improved solutions to the clutter rejection problem.

COMPUTER PROGRAM 1

```
      X1=0
      X2=0
      X3=0
      Y1=0
      Y2=0
      Y3=0
      Y = 0
100   IF(TEST(1).GT.0)GO TO 100
10    CALL ADDA (X,1,Y,1)
      Y = .923*Y1 - .645*Y2 + .025*Y3 + .452*X - 1.36*X1 + 1.36*X2 - .45
      X3 = X2
      X2 = X1
      X1 = X
      Y3 = Y2
      Y2 = Y1
      Y1 = Y
      GO TO 100
      END
```

```
      X1=0
      X2=0
      X3=0
      Y1=0
      Y2=0
      Y3=0
      Y = 0
100   IF(TEST(1).GT.0)GØ TØ 100
10    CALL ACDA (X,1,Y,1)
      Y = 1.94*Y1 - 1.37*Y2 + .3*Y3 + .805*X - 2.415*X1 + 2.415*X2 - .80
      X3 = X2
      X2 = X1
      X1 = X
      Y3 = Y2
      Y2 = Y1
      Y1 = Y
      GØ TØ 100
      END
```

METASYMBOL SUBROUTINE

For faster A/D - D/A conversion

```
X1      EQU     1
X2      EQU     2
A       EQU     5
B       EQU     4
$ADDA   PZE
        BRM     $RETURN
        PZE     4
AD      PZE
NAD     PZE
DA      PZE
NDA     PZE
        SKR     FINFLG
        BRU     $-1
        LDA     *NDA
        MRG     =077700000
        CØPY    (A,X1)
        BRX     $+2,1
        BRU     ATE0
        STZ     DCH
        BRU     $+2
RT8I    MPT     DA
        NOP     DCH
        LDP     *DA
        LLSB    15
        ARSB    15
        CØPY    0777,(B,X2),(C,B)
        CØPY    (X2,A),(A,X2)
        CØPY    (-A,A)
        CØPY    (X2,A),(A,X2)
        ARSD    0,2
        ETR     =077777000
        ADD     DCH
        STA     DACSM,1
        BRX     RT8I,1
        STZ     FINFLG
```

```
         LDA     BRMFIN
         STA     041
         LDA     *NDA
         LLSA    15
         ADD     =DACOM
         STA     DACW
         EOM     035001
         POT     DACW
AT9D     LDA     *AD
         STA     COUNT
         SKR     COUNT
         BRU     $+2
         BRR     ADDA
         LLSA    15
         ADD     =ADCOM
         STA     ADCW
         SKN     FINFLG
         BRU     $-1
         STZ     FINFLG
         LDA     BRMFIN
         STA     040
         EOM     034001
         POT     ADCW
         LDX     =00100000,1
         SKN     FINFLG
         BRU     $-1
IT5R     LDA     ADBUF,1
         COPY    (O,B)
         FLA     0.0
         STD     *AD
         MPT     AD
         SKR     COUNT
         BRX     IT5R,1
         BRR     ADDA
BRMFIN   BRM     FIN
```

```
FIN     PZE
        SKR     FINFLG
        BRU     $-1
        BRC     *FIN
FINFLG  PZE
DCW     PZE
CØUNT   PZE
CØN     FØRM    9,15
ADCW    PZE
ADCØM   CØN     0,ADBUF
        DATA    0,0,0,0,0,0,0,0,0,0,0,0,0,0,0
        DATA    0,0,0,0,0,0,0,0,0,0,0,00,0,0
        DATA    0,0,00,0,0,0,0,0,0,0,0,0,0,0,0
ADBUF   RES     32
DACW    PZE
DACØM   RES     13
        END
```

LIST OF REFERENCES

1. Nathanson, F. E. and Reilly, J. P., "Clutter Statistics which Affect Radar Performance Analysis," Supplement to the IEEE Transactions on Aerospace and Electronic Systems, v. AES-3, p. 386-398, November 1967.

2. Rome Air Development Center Technical Report RADC-TR-65-442, Microminiaturized MTI Group AN/GPA(XW-2), by W. L. Cappadona, p. 5-12, March 1966.

3. Roecker, W., "The Application of Digital Filters for Moving Target Indication," IEEE Transactions on Audio and Electro-acoustics, v. AU-19, p. 72-77, March 1971.

4. Linder, R. A. and Kutz, G. H., "Digital Moving Target Indicator," Supplement to IEEE Transactions on Aerospace and Electronic Systems, v. AES-3, p. 374-385, November 1967.

5. Skolnik, M. L., Radar Handbook, p. 35-1 - 35-17, McGraw-Hill, 1970.

6. Knowles, J. B. and Olcayto, E. M., "Coefficient Accuracy and Digital Filter Response," IEEE Transactions on Circuit Theory, v. CT-15, p. 31-41, March 1968.

7. Gold, B. and Rader, C. M., Digital Processing of Signals, McGraw-Hill, 1969.

8. White, W. D., "Synthesis of Comb Filters," Proceedings of the National Conference on Aeronautical Electronics, p. 279-285, May 1958.

9. Ruvin, A. E. and White, W. D., "Recent Advances in the Synthesis of Comb Filters," IRE National Convention Record, p. 186-199, March 1957.

10. Hoisington, D. B., Unpublished notes relating experimental work done on a solid state moving target indicator for the AN/UPS-1 radar, Naval Postgraduate School, January 1970.

11. Athey, S. W., Magnetic Tape Recording, National Aeronautics and Space Administration, January 1966.

12. Zverev, A. I., Handbook of Filter Synthesis, John Wiley, 1967.

13. Weinberg, L., Network Analysis and Synthesis, McGraw-Hill, 1962.

INITIAL DISTRIBUTION LIST

		No. Copies
1.	Defense Documenation Center Cameron Station Alexandria, Virginia 22314	2
2.	Library, Code 0212 Naval Postgraduate School Monterey, California 93940	2
3.	Professor D. B. Hoisington, Code 52Hs Department of Electrical Engineering Naval Postgraduate School Monterey, California 93940	1
4.	Lt. David L. Cummings, USN 400 S. Lewis Street Lakewood, Colorado 80226	1

Security Classification

DOCUMENT CONTROL DATA - R & D

(Security classification of title, body of abstract and indexing annotation must be entered when the overall report is classified)

1. ORIGINATING ACTIVITY *(Corporate author)*

Naval Postgraduate School
Monterey, California 93940

2a. REPORT SECURITY CLASSIFICATION

Unclassified

2b. GROUP

3. REPORT TITLE

An Application of Digital Filters for Radar Clutter Reduction

4. DESCRIPTIVE NOTES *(Type of report and, inclusive dates)*

Master's Thesis; March 1972

5. AUTHOR(S) *(First name, middle initial, last name)*

David Lee Cummings

6. REPORT DATE	7a. TOTAL NO. OF PAGES	7b. NO. OF REFS
March 1972	44	13

8a. CONTRACT OR GRANT NO.

b. PROJECT NO.

c.

d.

9a. ORIGINATOR'S REPORT NUMBER(S)

9b. OTHER REPORT NO(S) *(Any other numbers that may be assigned this report)*

10. DISTRIBUTION STATEMENT

Approved for public release; distribution unlimited.

11. SUPPLEMENTARY NOTES

12. SPONSORING MILITARY ACTIVITY

Naval Postgraduate School
Monterey, California 93940

13. ABSTRACT

This thesis presents the design procedures involved in implementing a digital filter for use in a range-gated moving-target-indicator (MTI) radar. The digital MTI filter synthesized on the XDS-9300 computer was a third-order recursive filter with Chebyshev characteristics.

The digital MTI filter was evaluated by constructing the equivalent of a single range channel for use with the AN/UPS-1 air-search radar and comparing its performance to that of the single delay-line canceler of the AN/UPS-1.

DD FORM 1 NOV 65 1473 (PAGE 1)

S/N 0101-807-6811

Security Classification

A-31408

KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
Digital Filters						
Moving Target Indicator						
Radar						